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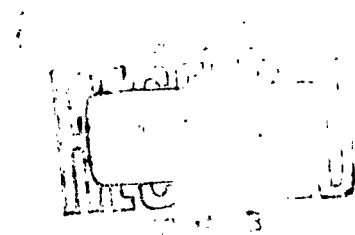
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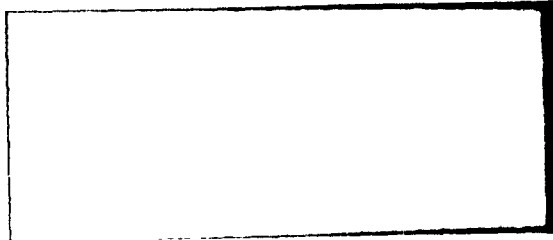


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ZERO FIELD-DISTORTION MICROWAVE
PROBES

by

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Contract Number AF30(602)-2336

Project Number 5561

Task Number 556103

Prepared
for

Rome Air Development Center
Research and Technology Division
Air Force Systems Command
United States Air Force

Griffiss Air Force Base
New York

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ABSTRACT

At least four types of probes are required to probe the high intensity microwave fields and the accompanying plasmas, which are the subjects of this investigation, with a minimum of disturbance due to the presence of the probes to the microwave and plasma properties. The disturbance of the plasma and the fields due to the volume displacement of the physical presence of the probe and stem, and due to the small amount of energy extraction necessary for the probe readings is, of course, unavoidable, but scattering due to the antenna-like configuration of the probe stem can be minimized to such a degree as to be negligible.

Precautions of design are necessary to insure that the disturbance due to the probe is kept near this unavoidable minimum. This unavoidable disturbance of the probes can be kept quite small and is essentially negligible because the power level of these experiments is many tens of DB higher than the internal noise level of the probe recording instruments, and therefore, the percentage of power required for the measurements and the actual probe volume can be such as to produce no consequential errors. The dimensions of the volume occupied by the sensitive elements of the probes are made small compared to a wavelength, and will therefore produce a minimum of disturbance to the microwaves.

The stems which support the probes and which contain the transmission line connections should be as small as mechanical and electrical designs will readily permit, and must be designed in a special way in order to minimize scattering of the microwaves.

Two well known methods for minimizing this scattering are available. The design of probe support and connections with absorbing surfaces is one method which is not suitable in this case because of the high powers which would have to be dissipated in the stem surface at the power levels required for these experiments. It is therefore necessary, and in any case more effective, to design a structure upon the surface of the probe stems which has a high wave impedance for the microwaves, compared to the impedance of space. This permits the currents excited in the probe stems by the microwaves to be held to a minimum so that their reradiation scattering will not interfere with the probe measurements or the general course of the experiment.

It has been found practical to design probes meeting these requirements for approximately 100 megacycles band width. Probes have so far been designed and tested which are essentially responsive to microwave magnetic field, microwave electric field, microwave power, as a function of wave direction, as in a directional coupler, and temperature of a small test object exposed to the microwave and to any plasma which may exist in its vicinity.

The methods of design, the designs, and results of tests are given in this technical note.

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INTRODUCTION

The experimental program required under Contract AF30(602)-2336 involves the measurement of microwave fields and the thermal state of test objectives in regions subject to high power microwave transmission and absorption and high power plasma maintenance.

These measurements will be accomplished by means of appropriate sensing elements mounted on the ends of long slender probes which must be inserted into the high intensity field and plasma regions with a minimum of disturbance to the wall-free microwave field and/or plasma concentration. These probes must be so designed as to withstand an appreciable incidental absorption from the microwaves and the plasma, which may produce a considerable temperature rise in the probes and the sensors. The probes and the supports which penetrate into the microwave fields and the plasma must therefore be constructed of materials which will maintain their physical and electrical integrity in spite of appreciable temperatures above ambient. These probes must be so designed as to absorb and scatter a minimum amount of energy from the fields and the plasma so as to produce the least disturbance of the conditions which will exist in free fields and plasmas and of the measurements from adjacent probes.

The frequency of the microwaves chosen for these experiments is 2350 megacycles. This is the highest frequency for which appropriate licensing was readily obtainable within the output range of the available Klystrons. The highest frequency available is desirable in order to produce the maximum microwave intensity in a free field with the available power. This frequency choice has its effect on all features of the probe design.

DESIGN OBJECTIVES

The design objectives for these probes can be easily discussed in two categories. General objectives, which apply to all the probes, and objectives applying to the individual probes which depend upon their particular intended use.

General Design Objectives - The general design objectives which apply to all probes depend upon mechanical and electrical requirements and combinations of these requirements. Mechanically, the probes must be stiff so that the positions of their sensors are readily determinable and reproducible. They must be smooth enough and stiff enough at their lower extremities so that they move readily through the sliding vacuum seals into the plastic sphere atmosphere enclosure. They must withstand fairly high temperatures up to perhaps a thousand degrees Fahrenheit or more at the sensing extremity. They must be at the same time as slender as is compatible with reasonable mechanical design in order to minimize the problems of undesirable energy pickup and microwave scattering. These requirements must be met with reasonable allowance for difficulties of mechanical construction.

Electrical Design Objectives - The general requirements of electrical design are that the transmission line in the interior of the probe stem must function with reasonable attenuation between 10 and 20 DB, or less, and with a reasonable standing wave ratio so that they can be connected to appropriate instrument loading with an overall standing wave ratio of the order 1.2 to 1.5, or less. Compatible with these limitations, the exterior parts of the probes must be designed so that the currents flowing up and down the supporting stem on the outside are low enough for a given wave excitation so that the scattered microwave power is 30 DB, or more, below the incident power. These objectives must be met with probes which are small in cross section compared to wavelength, as it is expected that the minimum size of areas of more or less uniform properties for both plasma and microwave field intensity will have an extent of about $1/4$ th wavelength, or approximately one inch. The probe size should therefore be such that compatible with other requirements it occupies a volume which is a small fraction of a cubic inch for each inch of linear length, along the portion of the stem which may extend into the strong field region during use. Figure 1 shows the electric and magnetic probes.

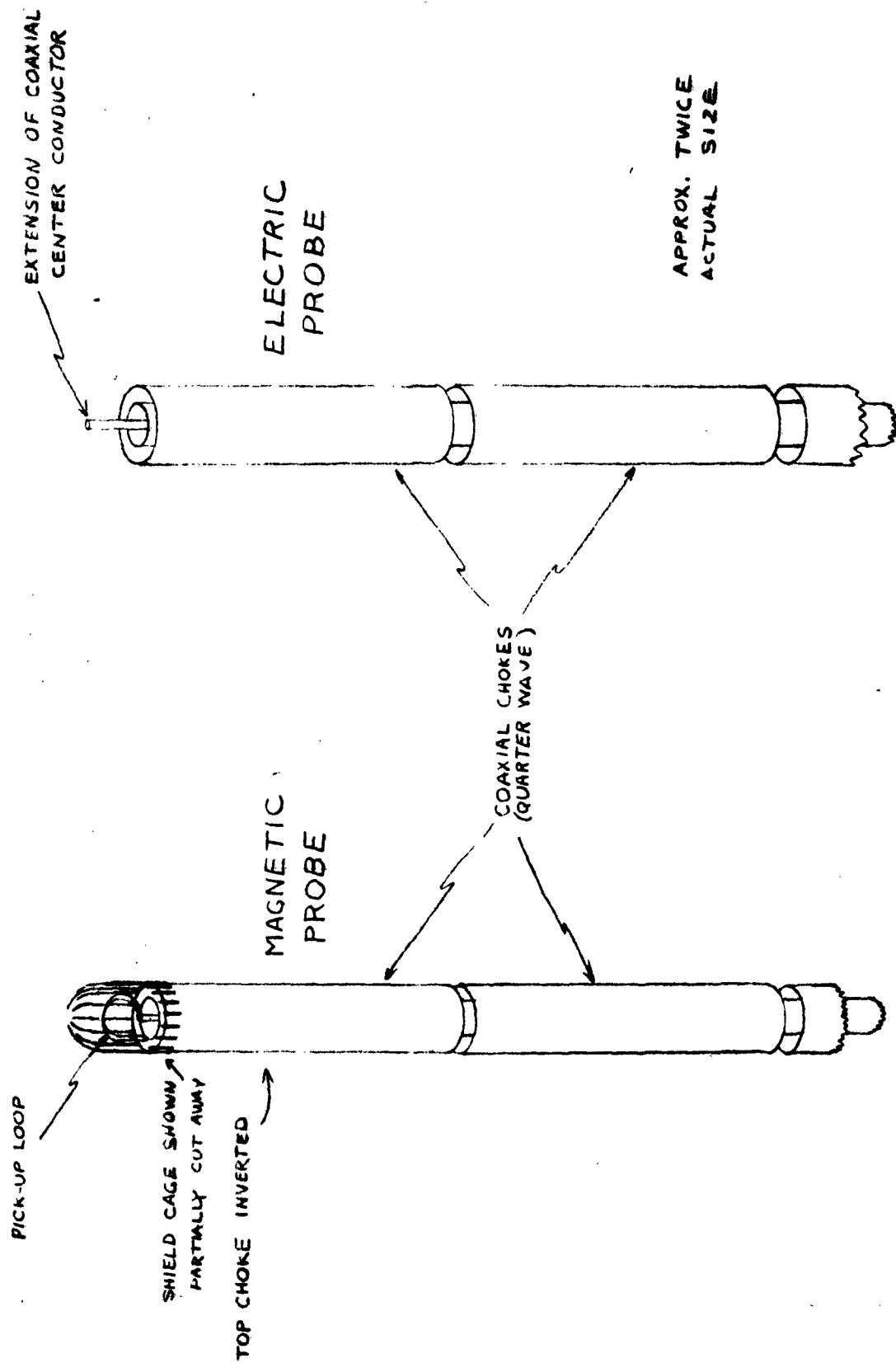


FIGURE 1. MICROWAVE FIELD PROBES

Individual Probe Design Objectives

Electric Field Probe - The Electric Field Probe will be expected to abstract a small amount of energy from the microwave field which is proportional to the electric field intensity of the wave in the immediate region where the probe is inserted, and which is proportional to the projection of the electric field vector on the probe axis direction. In a standing wave field where the magnetic field and electric field maxima are displaced from each other, it is expected that the response of this probe when placed at a magnetic field maximum will be 15 to 20 DB down from its response when placed at the electric field maximum for the same standing wave conditions. It is desirable to make some provision for at least a step adjustment of the extension of this probe so that the minimum disturbance to the field is introduced which is compatible with the requirements of the attached measuring instrument for fields. This probe must be mated to its nonreflecting supporting stem in such a way as to introduce a minimum of interference between the required properties of the stem and the sensing ability of the probe.

Magnetic Field Probe - The Magnetic Field Probe must respond to the magnetic field component of the microwave in a way similar to a magnetic coupling loop and must be shielded to a reasonable degree from response to electric fields. In a standing wave pattern, the response of this probe in electric field region of the standing wave should be at least 15 DB below its maximum response in the magnetic field region of the same standing wave pattern. This probe must be so mated to its stem that the nonreflecting properties of the stem do not interfere with its sensing function, and it must have an extent beyond the stem which minimizes the scattering due to the probe insofar as is compatible with a sufficient energy pickup to operate the associated readout instruments.

Directional Wave Probe - The directional wave probe will be expected to be able to respond separately to the wave energy propagated in two opposite directions, indicated by the orientation of the probe, and uniformly polarized along the probe stem axis. The distinguishability of the wave energy in two directions should be at least 15 DB, and preferably 20 or more DB. Such a probe will require two lossy transmission lines enclosed in its support stem, one for each connection to the instrument for reading the power in each of the two opposite wave directions. The requirements for mating this probe to the supporting stem and non-interference between the probe properties and the nonreflecting properties

of the stem are essentially the same as for the other probes. It is hoped that this probe can be made essentially independent of its position in any standing wave pattern which is encountered.

Thermal Probe - The thermal probe will consist of a nonreflecting support stem similar to that used for the other probes with provision for support of a test probe object and thermal contact between the test probe object and a calibrated thermocouple. The thermocouple lines must be provided with high impedance filters internal to their transmission line shield so that there is no appreciable heating of the thermocouple due to incidental microwave currents in the thermocouple or its leads, or be provided with sufficient shielding from the microwaves to serve the same purpose. The thermocouple must consist of a thermoelectric pair which is capable of withstanding high temperatures and operating as a thermocouple, with reasonable linearity between temperature and output voltage for the appropriate temperature.

DETAIL DESIGNS

General Stem Design - The mechanical and thermal requirements for a stem design require considerable mechanical strength at relatively high temperatures in a commonly available material. A good answer to this problem appears to be non-magnetic stainless steel which can be obtained in a variety of sizes of tubing and rod.

Fortunately, the probe length requirements and the permissible attenuation are compatible with the use of stainless steel provided the nonmagnetic requirement is adhered to so that skin depths for 2350 megacycles do not become excessively thin. Consideration of availability, reasonable ease of machining, stiffness, and reasonable volume displacement have been met by a choice of 1/8" outer diameter stainless steel alloy, having a resistivity of 72 micro-ohm centimeters, will be used throughout.

An innerconductor of stainless steel rod or tubing having 50 mils outer diameter is used with 1/8" tubing to form a coaxial transmission line having a characteristic impedance of 47.6 ohms. This innerconductor is .0007 inch too large for a standard 50 ohms and can be dressed down to the required diameter for the standard line impedance if a quarter wavelength matching section at the end of the probe proves to be inadequate in providing for a low standing wave ratio. It is believed, however, that in view of the attenuation and the accurate frequencies of operation, the matching section which will be built in to the transition from this line to a standard type N connector will be adequate. The series resistance on this line, based on a resistivity for the stainless steel of 72 micro-ohms centimeter at 2,350 megacycles, will be 9.1 ohms per foot. This will produce an attenuation of approximately 0.83 DB per foot, or in the neighborhood of 3-1/3 DB for the required 4 foot length of total probe support transmission line associated with any one probe.

The radio frequency resistance of the outside of this conductor is approximately 2.52 ohms per foot for uniform current distribution around the cylinder. This is approximately 1/2 an ohm per half wavelength, which is relatively negligible compared to the 72 ohm radiation resistance of a 1/2 wave dipole. This high ratio indicates that the effect of this external resistance on the reradiation scattering from a rod of this material will be negligible. Therefore, since the probe will be many wavelengths long, some special provision must be made to limit the circulating currents along the probe which are excited by the incident waves to a

sufficient degree to minimize scattering of the incident radiation and the interference and ambiguity which this scattering would produce in the measurements by this probe, and by the other associated probes.

One of the most effective methods of minimizing the radio frequency current flow along a rod which is exposed to radiation, is to place a series of re-entrant coaxial sections along the outside surface which are approximately a quarter wavelength long. This creates a series of resonant cavities which have a very high impedance to current flow and, therefore, allows a given electric field strength in the wave, to produce only small re-radiating currents. An idea of the effectiveness of this procedure can be obtained by comparing the resonant effective resistance of two such sections facing each other at their open ends, to the split half wave dipole point radiation resistance which represents their space coupling to an oncoming wave. The resonant section which has been designed for use by the magnetic field and electric field and thermal probe for this project consists of 1/4" outer diameter sleeves which are silver soldered at one end to the 1/8" outer diameter stainless steel tubing which constitutes the probe support and the outer conductor for the probe sensor transmission line. These sleeves have a 15 mil wall thickness so that their inner diameter is 220 mils. This forms a short section of transmission line shorted at one end to an inner conductor of 125 mils diameter. The characteristic impedance of such a section of transmission line is given by

$$Z_0 = 138 \log_{10} \frac{D_o}{D_i}$$

where D_o is the inner diameter of the outside conductor and D_i is the outer diameter of the inner conductor.

In this case, the value of Z_0 given by this equation is 33.88 ohms. The resistive impedance of such a quarter wavelength section short circuited at one end is given by $R = Z_0 Q$, where Q is defined in the usual way as 2π times the energy stored per cycle, divided by the energy lost per cycle. In a lumped circuit where the resistance and inductance of the circuit can be considered to be distinct and separate, this reduces to the usual equation

$$Q = \frac{\omega L}{R}$$

Since in this case both the inductance and the resistance are distributed, the usual calculation is required. The energy stored may be determined from

$$W = \int_{x=0}^{x=\frac{\lambda}{4}} \frac{1}{2} L_0 I^2 dx$$

where L_0 is the inductance per unit length, I is the current at each point x , and dx is the differential distance along the transmission line section. In the case of coaxial transmission lines, even with stainless steel with microwave frequencies, the skin depth is so thin that its thickness can be ignored in comparison with the other physical dimensions of the system. The effect of the inductive energy stored in the skin depths is to reduce the velocity of wave propagation slightly. This effect is, however, so small that it can be neglected and the propagation velocity can be taken as equal to the velocity of light in free space without a loss of accuracy sufficient to be of importance in this type of calculation.

We therefore assume a standing wave current on this distributed circuit such that the current

$$I = I_0 \cos \left(\frac{2\pi x}{\lambda} \right),$$

where the line is shorted at the point $x = 0$, and open at the end of the section where

$$x = \frac{\lambda}{4},$$

and the current is 0. The stored energy integral thus becomes

$$W = \int_{x=0}^{x=\frac{\lambda}{4}} \frac{1}{2} L_0 \left(I_0 \cos \frac{2\pi x}{\lambda} \right)^2 dx$$

and taking the constant terms outside the integral sign,

$$W = \frac{1}{2} L_0 I_0^2 \int_{x=0}^{x=\frac{\lambda}{4}} \cos^2 \frac{2\pi x}{\lambda} dx$$

Integrating, we have

$$W = \frac{1}{2} L_o I_o^2 \left[\frac{\lambda}{2\pi} \left(\frac{\pi x}{\lambda} + \frac{1}{4} \sin \frac{4\pi x}{\lambda} \right) \right]_{x=0}^{x=\frac{\lambda}{4}}$$

or

$$W = \frac{\lambda}{16} L_o I_o^2$$

Now, consider the total inductance of the system,

$$L_T = \frac{L_o \lambda}{4},$$

$$W = \frac{1}{2} L_T I_o^2,$$

which is one-half the value of stored energy which would appear on the lumped circuit with the same total inductance.

The losses in the system can be approached in the same way and will be given by the equation

$$H = \int_{x=0}^{x=\frac{\lambda}{4}} R_o I^2 dx$$

substituting for I as before

$$H = R_o I_o^2 \int_{x=0}^{x=\frac{\lambda}{4}} \cos^2 \frac{2\pi x}{\lambda} dx$$

or,

$$H = \frac{R_o I_o^2 \lambda}{8}$$

Now, consider that the total resistance $R_T = \frac{R_o \lambda}{4},$

we have $H = \frac{R_T I_o^2}{2},$ which is just one-half the loss which would occur in a lumped circuit of the total resistance.

Thus, the Q can be determined by the equation $Q = \frac{\omega L_T}{R_T}$, or since

$$L_T = L_o \frac{\lambda}{4}, \text{ and } R_T = R_o \frac{\lambda}{4}, \quad Q = \frac{\omega L_o}{R_o}.$$

It is convenient to calculate L_o in terms of the Z_o of the line, taking advantage of the fact that the wave velocity will approximate free space velocity with sufficient accuracy. As is well known, the velocity c in a transmission line is

$$c = \frac{1}{\sqrt{L_o C_o}}$$

and the characteristic impedance of a line Z_o equals $\sqrt{\frac{L_o}{C_o}}$.

Substituting the value of the $\sqrt{C_o}$ from the velocity equation in the impedance equation gives

$$\frac{Z_o}{c} = L_o$$

Substituting in this equation for Q, we have

$$Q = \frac{2\pi f Z_o}{c R_o}$$

Since $F\lambda = c$, this gives Q in terms of Z_o

$$Q = \frac{2\pi}{\lambda} \frac{Z_o}{R_o}.$$

It is convenient to calculate R_o from an approximate equation well known to radio engineers. This formula gives the radio frequency resistance of a cylindrical copper conductor having uniform current distribution as $R_o = \sqrt{\frac{f}{D}}$ ohms per thousand feet, where F is in megacycles and D is the diameter of the conductor in inches. This formula is readily modified for other materials which are non-magnetic by multiplying the right hand side of the equation by the square root of the ratio of the resistivity of the material in question to the resistivity of copper.

$$\text{Thus, } R_o = \sqrt{\frac{2350 \text{ mc} \times 72 \text{ micro-ohms cms}}{1.7 \text{ micro-ohms cms}}} \frac{1}{D \text{ inches}} \quad \text{ohms per thousand feet}$$

for stainless steel type 304.

This can be reduced to ohms per centimeters by dividing by the factor 30.48 centimeters per foot.

$$\text{So, } R_o = \frac{315.48}{30.48 D} 10^{-3} \text{ ohms per centimeter for each conductor.}$$

The quarter wavelength section under consideration here will have current in both the inner and outer conductor so that the resistance of the two conductors must be added.

Since the microwave frequency resistance of these conductors varies inversely with the diameter, the resistances of the inner and outer conductors must be calculated separately and added. The resistance per centimeter of the outer conductor for this resonator is the

$$R_{oo} = \frac{10.350 \times 10^{-3}}{0.220} = 0.0470 \text{ ohms per centimeter,}$$

and for the inner conductor

$$R_{oi} = \frac{10.350 \times 10^{-3}}{0.125} = 0.828 \text{ ohms per cm, and}$$

$$R_{oo} + R_{oi} = 0.470 + 0.0828 = 0.1298 \text{ ohms per cm} = R_o.$$

Substituting this value in the equation $Q = \frac{2\pi Z_o}{R_o \lambda}$

One wavelength at 2350 = 12.76 inches, this gives

$$Q = \frac{6.2832 \times 33.88}{0.1298 \times 12.76} = \frac{212.87}{1.6562} = 128.5.$$

The shunt impedance of this cavity of resonance is then given by

$$Z_R = QZ_o = 128.5 \times 33.88 = 4353.58 \text{ ohms.}$$

Since two of these cavities will be face to face to compare to a half wave center feed dipole of 72 ohms radiation resistance, the current ratio between this system and a smooth dipole will be

$$\frac{2 \times 4353.58}{72} = 120.93.$$

The power scattered will be proportional to the square of this ratio, or 14624. This is a ratio of 41.6 DB and represents the attenuation of scattering which may be expected for the proposed choke system on the outside of a probe stem at its optimum rejection frequency.

The band width for which this anti-scattering method is effective can be estimated by referring to the universal resonance curves given in "Radio Engineering"¹. Here we see that for a frequency deviation for resonance of

$$\frac{2 f_o}{Q}$$

the impedance of the resonance circuit will be approximately 25% of its resonant value at the f_o frequency. This is a deviation for our design frequency of 2350 megacycles, of

$$\frac{2350}{70.5}$$

or 33-1/3 megacycles. At this frequency the impedance will be

$$\frac{8707}{4} \text{ or } 2177 \text{ ohms.}$$

Referring to the half-wave dipole curves shown in "Antennas - Theory and Practice"², 1-1/2% off resonance gives a radiation resistance of approximately 60 ohms and a capacity reactance of approximately 20 ohms which for this purpose is negligible.

The ratio of 2177 to 60 is 36.3 and the square of this ratio is 1317 which still represents a rejection of approximately 31.2 decibels sufficient to meet our scattering requirements of 30 decibels rejection.

On the other side of the resonance curve at 33-1/3 megacycles higher frequency, we have the same resonance impedance, but the antenna resistance is up to about 80 ohms which proceeding as before gives a ratio of 27.2. The square of this is 740 which represents a rejection ratio of 28.7 decibels which is quite close to the required 30 decibels rejection. Allowing for this discrepancy we may say that

1. F. E. Terman, Radio Engineering (2nd Edition, McGraw Hill Publishing Co.) p. 56.
2. Sergei A. Schelkunoff & Harald T. Friis, Antennas - Theory & Practice, (John Wiley & Sons, 1952 Edition), p. 449.

the satisfactory band width for rejection of this arrangement is approximately 65 megacycles shifted slightly to the low frequency side of the resonance frequency of the quarter wavelength choked sections. This is quite satisfactory rejection band width for the fixed frequency probing for which these probes are designed.

The wavelength appropriate to these designs is very close to five inches. One-fourth of this wavelength is therefore 1.250 inches. The inner length of the rejecting sections on the outer conductor should have an electrical length of one-fourth wavelength. However, these sections produce the best rejection when placed with one-eighth inch separation. This separation and the fringing fields at the end of the outer conductor of each rejection section loads the resonance capacity so that the inner physical length must be foreshortened. This foreshortening is difficult to calculate because of the complex mathematics involved in predicting fringing fields. Experimentally it is found that a foreshortening of about 12% is satisfactory. The sections are therefore constructed with an internal length of 1.100 inches which checks experimentally for minimum scattering at 2350 megacycles.

Similar calculations can be made for the larger diameter probe which will be used for directional wave detection. The larger size will allow for higher Q in the rejection sections and therefore a better rejection at the optimum design frequency. The band width of this probe will not be exactly the same but the unavoidable displacement due to its physical volume will of course be greater.

These estimates will be checked by scattering measurements made by the probes in the actual microwave field waves under test.

Probe Sensors

Electric Field Probe - The Electric Field Probe will consist merely of an extended central conductor beyond the end of the coaxial line and its non-reflecting system. Any length less than $1/8$ wavelength should induce very little scattering and will be satisfactory provided it picks up sufficient energy to meet the power requirements of the indicating meters.

Experiments with such a probe one centimeter long operated at 2350 megacycles or 12.76 centimeters wavelength gives satisfactory results at a 1 milliwatt level. Such a probe should pick up approximately 27 volts from a field of one watt per

square centimeter. The impedance of such a probe as given by "Antennas - Theory and Practice"³, should have a radiation resistance of about 6 ohms, and a reactance of approximately 80 ohms. The radiation resistance can be in this case neglected in determining the proportion of the voltage which will appear across the 47.6 ohm probe transmission line. The part of the 27 volts which will appear across the line is given by 27.6 divided by the absolute impedance of the combined impedance of the probe and transmission line, or 47.6 divided by the square root of $47.6^2 + 80^2$, or 51%; that is, approximately 14 volts peak should appear across the input end of the probe line of 1 watt per square centimeter power density.

The probe power which enters the probe line is $\frac{E_o^2}{2 Z_o} = \frac{(14)^2}{2 \times 47.6}$

This corresponds to a probe cross section of 2.1 square centimeters. This power is reduced by 3.3 DB by the probe transmission line so that 0.975 watts is available to the metering connection. This is obviously too much pickup to be used in our maximum field strength of 500 volts per centimeter, or about 330 watts per square centimeter.

A probe length of approximately only 1 millimeter will be used for the high fields. The 1 centimeter probe length, however, is useful in comparing calculations as a 1 millimeter probe will be almost entirely in the fringing field of the end of the transmission line. This size probe should have 0.02 sq. centimeters cross section and should produce 9.75 milliwatts for metering.

The final probe sensitivity will be determined by calibration in fields, the strengths of which are known from other methods, so that comparison of the calibrations with calculations for this type of probe is only useful as a guide. Much smaller and less disturbing probes can be designed following the foregoing principles if more accurate measurements prove desirable or if the frequency is shifted to take advantage of the greater concentration which can be produced with shorter wavelengths.

Magnetic Field Probe - The magnetic field probe consists essentially of a small coupling loop connected across the probe stem transmission line. Such a

3. Sergel A. Schelkunoff & Harald T. Friis, Antennas - Theory and Practice, (John Wiley & Sons, 1952 Edition), p. 444.

loop will also respond to the electric field because of its high inductive impedance compared to the characteristic impedance of the transmission line. Some provision is therefore required to reduce the electric field response to a negligible value.

This can be done by a faraday shield of conducting spikes or pickets surrounding the coupling loop provided the whole assemblage is small enough compared to a wavelength so that phase shift along the pickets can be neglected. Pickets one-half centimeter long satisfy this requirement at 12.76 centimeters wavelength. A coupling loop of approximately square active area of 0.18 square centimeters surrounded by a cage of small conducting pickets operates with a satisfactory ratio of magnetic to electric field pickup of 40 DB or better as shown by measurements and comparison with an electric field probe previously described operating in a high ratio standing wave field.

The voltage response of such a probe in a radiation power field of one watt per square centimeter can be calculated for comparison with the electric field probe response. The voltage induced in the coupling loop will be proportional to the area of the loop which is presented perpendicular to the magnetic flux lines. The power available from this voltage will be proportional to the square of the area of the loop.

The peak voltage V_o generated in the loop is given by $V_o = \omega B_o A$ where B_o is the peak flux density, ω is the angular frequency of the field, and A is the area of the loop in square centimeters. B_o can be determined from the equation

$$\frac{Z_o H_o^2}{2} = S$$

where Z_o is the impedance of space, H_o is the magnetic intensity density in amperes per centimeter, S is the power density in watts per square centimeter, and the factor 2 is required because H_o is the peak value of a sine wave time variation of current density.

Recalling that

$$B_o = \mu_o H_o$$

we have then

$$B_o = \mu_o \sqrt{\frac{2S}{Z_o}}$$

Taking $S = 1$ watt per square centimeter and $Z_o = 377$ ohms, we have

$$B_o = \left(4\pi \times 10^{-9} \frac{\text{hen}}{\text{cm}} \right) \sqrt{\frac{2}{376.73 \text{ ohms}}} = 9.15 \times 10^{-10} \frac{\text{volt seconds}}{\text{cm}^2}$$

Substituting this value into the proper equation, we have

$$V_o = 2\pi f B_o A$$

$$V_o = 2\pi \times 2350 \times 10^6 \times 9.15 \times 10^{-10} \times 0.18$$

and $V_o = 2.4$ volts peak induced in this coupling loop (area = 0.18 cm^2) placed at optimum orientation in a one watt per square centimeter field at 2350 megacycles field frequency.

The inductance of this coupling loop can be calculated approximately from its geometry by means of the formula given in Bureau of Standards Circular No. 74⁴

$$L = \left\{ 9.21 \left[(a + a_1) \log_{10} \frac{2aa_1}{b+c} - a \log_{10} (a+g) - a_1 \log_{10} (a_1+g) \right] + 4 \left[2g - \frac{a+a_1}{2} + 0.441(b+c) \right] \right\} \times 10^{-9} \text{ henrys}$$

$$g = \sqrt{a^2 + a_1^2}$$

$$a = 0.5 \text{ cms}$$

$$a_1 = 0.36 \text{ cms}$$

$$c = 0.10 \text{ cms}$$

$$b = 0.02 \text{ cms}$$

This gives $L = 7.1 \times 10^{-9}$ henrys. The impedance of this loop at 2350 megacycles is given by

$$Z_1 = j\omega L = j(2\pi \times 2350 \times 10^6 \times 7.1 \times 10^{-9}), \text{ or}$$

$$Z_1 = 104.8 \text{ ohms.}$$

4. Bureau of Standards Circular No. 74 (2nd Edition) March 10, 1924

This will permit 41-1/3% of the voltage generated in the loop to appear across the transmission line or about 1 volt. The power input to the line will then be

$$W = \frac{(1)^2}{2 \times 47.6} = \frac{1}{95.2} = 0.0105 \text{ watts}$$

or 10.5 milliwatts from a wave field of one watt per square centimeter. This loop therefore has an equivalent cross section of 0.01 square centimeter as a wave energy collector.

The 3.3 DB loss in the probe stem transmission line leaves 4.9 milliwatts available to operate the readout meter system.

Directional Wave Probe - The Directional Wave Probe works on the same principal as that used in making directional wave couplers to wave guides and transmission lines. It has been found that these probes can be applied successfully in free space provided the two transmission lines which are necessary for the two wave directions are properly shielded from causing excess scattering in the wave field and provided that either sufficient attenuation is provided in the coupling lines or they feed into a matched load of sufficient accuracy.

This latter requirement is obviously necessary in order to prevent remixing of the wave directions after the separation which is produced by the probe itself.

The requirement for two transmission lines has made it desirable to use the same transmission line structure as in the other probes, with a nonscattering shield supporting a pair of these lines. This requires a larger overall probe structure. Two .125" outer diameter, outer conductor, transmission lines are inserted in stainless steel tubing having an outer diameter of .310" and a .015" wall. This allows a small amount of extra space so that two transmission lines can be mechanically assembled into the anti-scattering shield without undue mechanical problems.

The 1/4 wave length rejection sections are assembled on the outside of the .310 outer diameter tubing. They consist of sections of stainless steel tubing 500 mils outer diameter and 450 mils inner diameter. Their length appropriate to the 2350 megacycle design frequency is 1.100" internal length and 1.225" external length. A space of .125" between adjacent ends of these sections gives the appropriate capacity loading for these lengths and sizes.

The details of this assembly are shown in Figure 2. A special fitting at the lower end of this probe assembly shields the probe against vacuum leaks when it is used in our special atmosphere enclosure. This fitting provides the transition to two Type N connectors which are an integral part of the probe and which provide for connecting the two separate wave direction lines to appropriate matched loads or lengths of lossy transmission line, and to the appropriate meters for indicating the wave strength in the two directions, or the standing wave ratio at the vicinity of the probe.

The sensing element of the probe itself consists of a single wide flat coupling loop, the ends of which are connected to the interior conductors of the two transmission lines. Provision is made for balancing the electric field pickup of the two halves of the loop against their magnetic field pickup in such a way that the phase of additions and subtractions depends on the direction of the wave, and waves of one direction will be transmitted in one of the transmission lines, and in the opposite direction in the other. The final adjustment is made experimentally in a wave field with the standing waves being adjusted for a minimum of energy in the appropriate transmission line and then reversing the probe and adjusting for a minimum in the other. This procedure can be done in a converging series with an accuracy of balance of at least 30 DB. The attenuation of these transmission lines will be approximately 0.83 decibels per foot, or about 3-1/3 decibels in the 4' length of each probe line. This is not sufficient in itself as at least 15 DB loss is required in each line in order to make possible a 30 DB balance.

Additional lossy line can be added to make up this 15 DB, or accurate matching loads can be connected across the line. Both methods have been satisfactorily used. The small amount of energy required to operate the indicating meters can be extracted without upsetting the necessary load match.

The balancing of this probe in order to separate the responses to the two wave directions can be accomplished in two ways. A faraday cage of small stiff pickets similar to that used for an electric field shield on the magnetic field probe can be installed around the directional probe so that it is movable along the probe stem direction. This allows the electric field intensity to be cut down to the balancing intensity without appreciably upsetting the phase balance. Such a balancing device has proven practical from an electrical standpoint but is awkward to adjust mechanically.

An alternate method is to shunt the center of the loop with an adjustable capacitance formed between the end of a small screw and the cap through which the connecting transmission lines terminate next to the probe sensor. This arrangement also gives a satisfactory electrical balance and has the advantage of a continuous screw adjustment.

The latter method will be used unless thermal requirements cause unforeseen difficulties with the stainless steel adjusting screw. Sensitivity very similar to the magnetic field probe is obtained.

20 DB back to front ratio is an easy adjustment and 30 DB can be obtained with care in a field sufficiently free from reflections. A better field environment may make possible as much as 40 DB directivity. This will be investigated later on the outdoor range.

Thermal Probe - The Thermal Probe will consist of a stem exactly similar in external arrangements to the electric field probe and the magnetic field probe. The inner conductor of the transmission line will, however, be replaced by two insulated wires of appropriate thermoelectric materials. These will form a thermo-junction at the tip which will be in thermal contact with the sealed end of the probe transmission line outer conductor.

This outer conductor will extend far enough so that the thermal loss due to conductivity of the thin walls of the stainless steel will be small and reasonably constant.

Provision will be made to weld a sample test material to the tip of this outer conductor for metallic samples or to insert this tip into intimate thermal contact with dielectric samples. This method of closure, which depends on the low heat conductivity of the 0.006" stainless steel, will provide adequate shielding of the thermocouple from direct microwave currents. This thermocouple unit can be readily tested by comparison with standard thermometers or thermocouples, and with electrical power inputs supplied from 60 cycle sources so that temperature rise rates can be determined for known applied powers. The opposing couple must be completed at the output end of the probe where its temperature can be known and controlled, as all thermoelectric couples operate on temperature differences. Copper conductors to the indicating meter can only be used beyond

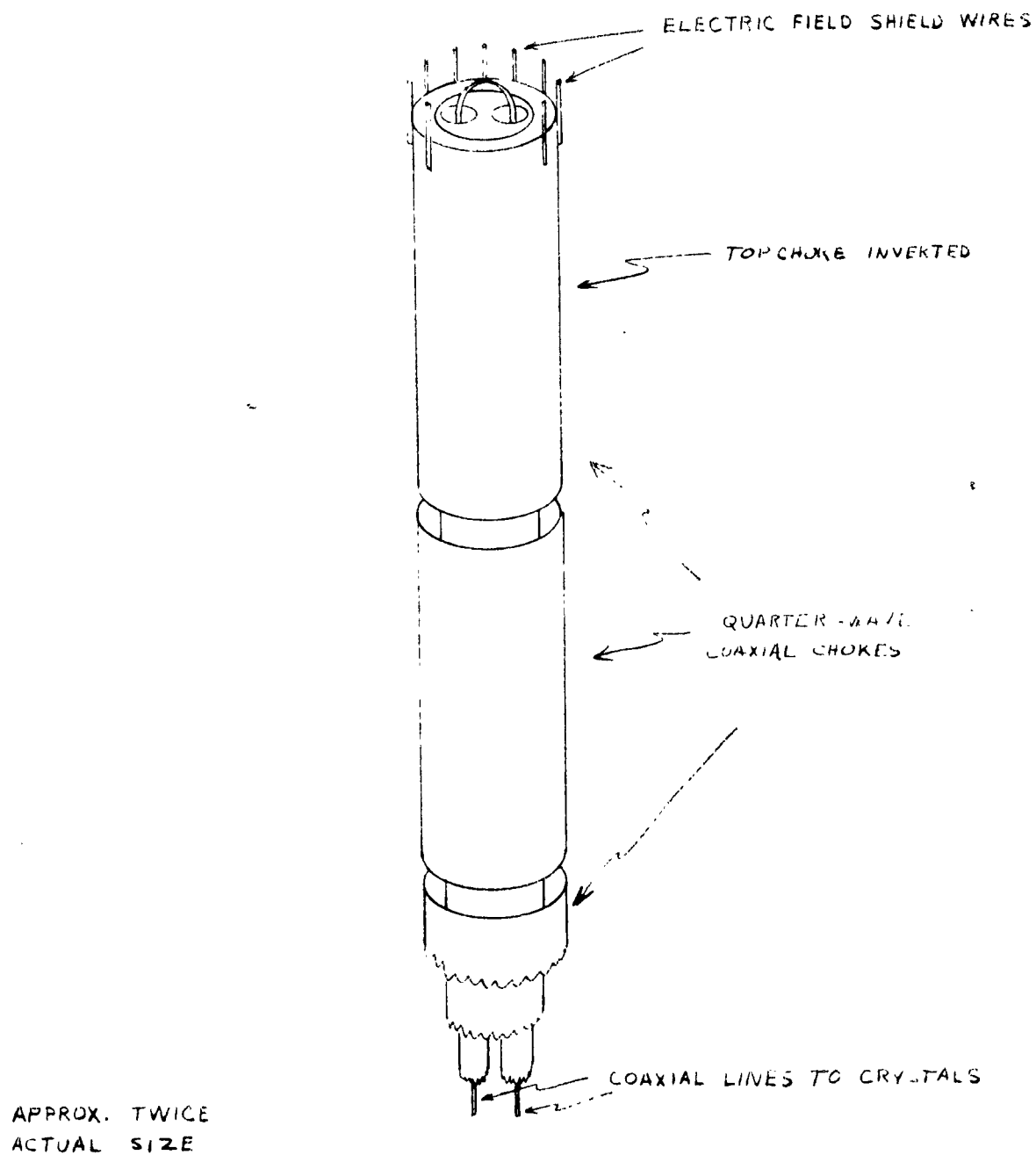


FIGURE 2. DIRECTIONAL PROBE

this comparison junction. A chromel-alumel couple will be investigated, as the temperature calibration for this couple is standardized to temperatures in the expected range.

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1. F. E. Terman, Radio Engineering (2nd Edition, McGraw-Hill Publishing Co.), p. 56
2. S. A. Schelkunoff and H. T. Friis, Antennas-Theory and Practice, (John Wiley & Sons, 1952), p. 449
3. Bureau of Standards Circular No. 74, (2nd Edition), March 10, 1924
4. Design of a Parallel Plate Microwave Lens, RADC TDR No. 62-431 (ASTIA AD-285 513).

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APPENDIX I
CALIBRATION OF ELECTRIC AND MAGNETIC PROBES

In order to obtain a useful quantitative calibration factor for the R.F. field probes, the concept of "receiving cross section" is used.¹ That is, if the sensitive region of a probe is placed in a field whose intensity is S watts/cm², and a power P is measured at the output connector of the probe, then the effective area A_r of the probe is

$$A_r = \frac{P}{S} \text{ cm}^2 \quad (1)$$

when the probe is feeding a matched termination (50 ohms in this case). After the effective area of a probe is determined in a weak but calculable field, the probe can be used with a suitable calibrated attenuator to quantitatively determine the intensity of a much stronger field.

Figure 3 shows the apparatus used to calibrate the probes. With this arrangement it is not necessary to have an instrument capable of measuring absolute R.F. power level (such as a bolometer), since only relative powers are measured, and this is done by re-establishing a previously measured power level by means of the calibrated attenuator. Hence unknown non-linearities of the detector crystal do not complicate the measurement.

Here power from a Hewlett-Packard Model 616B UHF oscillator (2350 mc) is fed to a standard gain horn via a calibrated attenuator and a wave-guide directional coupler (Tyne N connector output). Provided that the probe is placed in the "far field" of the horn, the field intensity at the probe is

$$S = \frac{P_o G}{4\pi R^2} \text{ watts/cm}^2 \quad (2)$$

where P_o = power leaving horn (watts)

G = gain of the horn

R = distance from mouth of horn to probe (cm)

The "far field" of the horn is the region of space where the field intensity falls off as R^{-2} , the inverse square of the distance from the horn.² This region is considered to begin at a distance D^2/λ , where D is the maximum dimension of the horn opening. For this case D is about 15 inches, and $\lambda = 5$ inches, so that the far field begins at about 45 inches. A distance of 2 meters was chosen for R so as to make computations easier. The horn, aimed upward, was positioned about 2 meters away from the wall of the building so as to be well outside the 26° beam width. The absence of reflections was verified by moving the electric probe vertically and observing no maxima or minima.

Upon substituting Eq. (2) into Eq. (1), we get

$$A_r = \frac{P}{P_o} \frac{4\pi R^2}{G} \quad (3)$$

$$= \frac{1}{K} \frac{4\pi R^2}{G} \quad (4)$$

where K is the ratio of the power in the waveguide to the power obtained from the probe. This ratio K is experimentally determined in decibels from the attenuator setting. A 5 db attenuator is always used ahead of the crystal to provide a good 50 ohm resistive termination.

If insufficient power is available to get noise-free readings, a VA-800 klystron can be operated at low power as a source of microwave power.

Calibration Procedure

The calibrated attenuator is set to some known attenuation (preferably zero), the attenuator-crystal assembly is connected to the probe, and a reading is obtained on the tuned voltmeter.

Then the attenuator-crystal assembly is transferred to the 43 db calibrated attenuator and sufficient additional attenuation is introduced by the calibrated attenuator to reduce the indication on the tuned voltmeter to that obtained previously. Thus the ratio K expressed in decibels is 43 plus the attenuation indicated on the calibrated attenuator, and the area A_r can be calculated from Eq. (4).

PROBE BEING
CALIBRATED

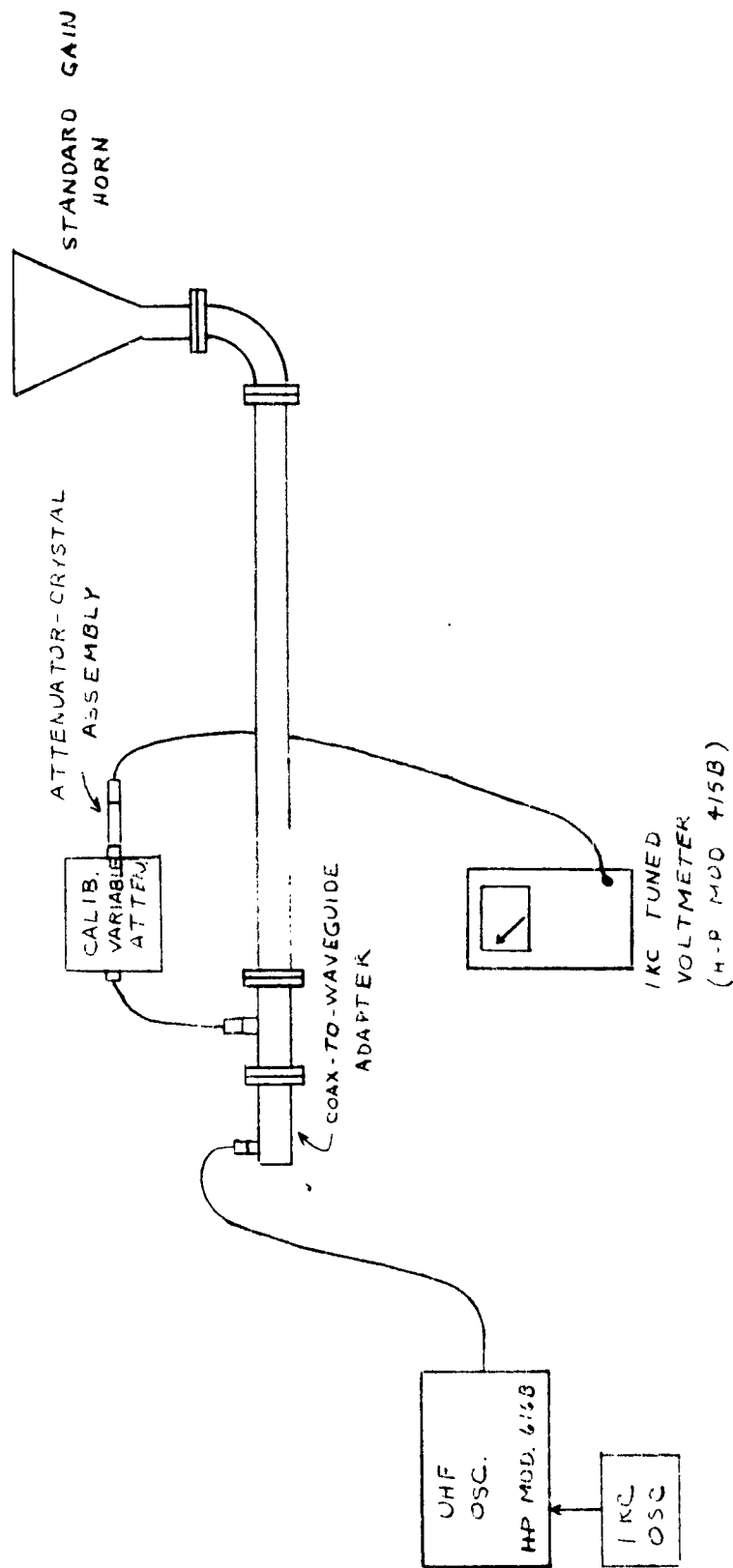


FIGURE 3. PROBE CALIBRATION APPARATUS

Calibration Results

Gain of horn at 2350 mc is 16.6 db (measured by manufacturer), equivalent to a power gain $G = 45.7$.

Electric Probe: The attenuation factor K as determined by the method described above is 54 db, so that $K = 2.5 \times 10^5$.

$$\begin{aligned} A_r &= \frac{1}{2.5 \times 10^5} \frac{4\pi (2 \text{ meters})^2}{45.7} \\ &= .44 \times 10^{-5} \text{ meter}^2 \\ &= 0.044 \text{ cm}^2 \end{aligned}$$

Magnetic Probe: The value for K here is 59 db, or $K = 8 \times 10^5$

$$\begin{aligned} A_r &= \frac{1}{8 \times 10^5} \frac{4\pi (2 \text{ meters})^2}{45.7} \\ &= 0.014 \text{ cm}^2 \end{aligned}$$